1. DAB interference to host FM signal.
2. Interference to DAB from the first adjacent FM signal.
3. Interference to the FM signal from the first adjacent DAB.
4. Interference between DAB second adjacents.
5. Robustness of DAB in multipath fading environment."

In reviewing the available literature for these IBOC DAB systems, it became clear that, while extensive testing has been done of proposed IBOC systems, no detailed modeling effort has been performed independently in an attempt to quantify the problem areas above. In this report, Mobile Data Systems has developed a Matlab simulation of the FM-1 system, both original and proposed, in an attempt to evaluate the tradeoffs in system design. Modeling complex systems instead of designing and building hardware is an effective and important technique for evaluating high risk technology without the enormous expense of creating and testing hardware. It is somewhat surprising that a detailed simulation was not performed on the USADR system; even this modest effort does not include a complete simulation of the analog FM signal.

B. Modeling Issues

In order to model the FM-1 system, the design parameters must be known or inferred; while details of the original FM-1 proposal are well known, details on the proposed modifications are not as well documented in [13]. In fact, [13] also includes a brief description of an orthogonal frequency division multiplexing (OFDM) system for DAB that appears to be completely unrelated to the DSSS system; in this report, we will only model the modified DSSS system. The modified system is, at this time, only a proposal.

The system parameters, whether gathered from documentation, derived, or inferred, are listed in Appendix C. From these parameters, both versions of the FM-1 system were simulated, to include bit error rate (BER) as a function of the energy per bit (E_b) divided by the noise power spectral density (N_0). This ratio, E_b / N_0, is similar to signal to noise ratio commonly used in analog systems. In this case, N_0 can include effects such as adjacent DAB channel interference and other sources of "noise like" interference. The effect of BER should be obvious; as the number of bit errors increases, the performance of the system completely breaks down. It is characteristic of digital systems (such as digital cellular) to have a "wall" or critical BER at which the system becomes unusable. Thus, unlike an analog system where the user will hear a gradual degradation of the signal, a digital audio system will degrade rapidly and fail to operate.

It should be noted that models such as this can be most effectively used to compare the performance of the system to the theoretical limits; the simulation output will establish whether the system can work in the ideal case, but the results of testing with prototype systems will not be explicitly compared here; the reader is referred to [14], which is a comprehensive report of laboratory testing. The models can effectively establish whether a system could work under ideal conditions; since real world systems do not consist of ideal components, these simulations establish upper bounds on performance.

C. Comments and Conclusion

The issues addressed in the modeling efforts allow us to draw conclusions about the performance of the USADR systems, both the original and modified versions. In general, the conclusions are that the modified USADR system corrects some deficiencies in the original design, but worsens its performance in several critical areas, leading to the conclusion that the adjustment of system parameters is a technical "shell game"; the USADR system operates at the limits of direct sequence spread spectrum performance, and attempts to optimize one parameter will likely cause degradation in one or more other parameters. Only significant reductions in source coding rate can lead to any real improvement in system performance; apparently, USADR maintains the same program data rate in both system proposals.

One of the most significant findings in this study is that the modified USADR system has considerably less processing gain than the original system (23dB in original vs 11dB in modified), which makes it much more susceptible to multipath, co-channel interference from the analog signal, from itself, and from adjacent channel interference. Processing gain is a measure of the ability of a DSSS system to operate in the presence of interference, and higher is better. In the case of the modified system, since the processing gain is only 11 dB, the simulation shows a rather alarming bit error rate of 0.04 for the modified system, with no analog interference present. Thus, while the modified system fits within a narrower spectral mask, USADR had to lower the processing gain to do so.

We shall address several specific areas in light of the simulation results:

1. Host compatibility

In the modified system, the spreading bandwidth and processing gain was reduced, and the injection level was lowered 6-10dB in order to reduce the likelihood of interference to the analog program. Indeed, the modified system will certainly be less likely to cause interference, but the IBOC signal is much more fragile, and requires a far more complicated receiver.[1] Thus, the modified system is better in terms of digital interference on the analog host, but worse in interference to the DAB system.

2. Multipath

Of particular interest is the multipath performance of the modified USADR system, particularly in light of the reduced processing gain. As outlined in Appendix D, the USADR modified operates at a marginal 0.04 BER in the <u>absence</u> of multipath due to self-interference. If multipath causes an additional degradation of 3 dB in E_b / N_0, then the

[1] See Appendix C; the modified system will require 512 correlators running in parallel, which is over 5 times more complex than the original FM-1 system.

BER will further decrease to approximately 0.06, which will be extremely difficult to overcome using error correction techniques. One possible technique to improve this is use of a RAKE receiver, as described in Appendix C; the coherence bandwidth[2] of fades in the FM band is within the range of the FM-1 bandwidth, which means that a RAKE receiver may not be useful for improving the performance of FM-1 in multipath; RAKE receivers cannot improve the signal if is completely wiped out by a fade, and the modified system will gain less from a RAKE receiver than the original system. The only techniques that will be useful for a relatively narrow signal such as this are frequency or spatial diversity. USADR employs frequency diversity by selecting either upper or lower sideband, which gives an overall E_b / N_0 improvement related to the statistical independence of the sidebands.

II. Appendices

Mobile Data Systems has designed a mathematical/functional system to match the original USADR FM-1 Noise Modulation and Coding (NOMAC) system, and FM-1 enhanced system based upon Gold codes. The assumptions we have used to develop computer simulation results for both systems will be clearly stated. We designed a canonical functional description in order that theoretical performance bounds are readily available; we compare simulated results to theoretical bounds to ensure the simulation is accurate, thus achieving independence from details of specific hardware implementation unavailable to MDS.

Computer simulations predict a degraded performance of the 48 channel NOMAC FM-1 system due to loss of signalling orthogonality over finite length waveforms, and a CDMA-like self interference due to the overlay of 48 channels to increase the channel bit rate. Details of signal filtering to form the baseband spectrum are unknown.

A 32 channel, 32-ary, 64 bit Gold code based enhanced FM-1 system was simulated in order to estimate the baseband spectral content of the waveform suite and predict interference effects; raised cosine baseband filtering, and 32-ary biorthogonal signalling are assumed. The extent of RF filtering is unknown. Additionally, computer simulations were run to predict performance degradation from the self interference of a 32 signal overlay that agree with large sample statistics CDMA BER estimates. The simulations indicate that baseband pulse filtering is critical to creating a digital sideband IBOC signal that meets FCC spectral mask requirements. The receiver complexity for the proposed enhanced system is high, requiring 512 correlators, each of which is a length matched to the chosen Gold code length.

[2] The coherence bandwidth of a radio channel is the bandwidth over which the statistical properties are highly correlated. It is related to the "delay spread" in the channel, which is the amount of time that reflected signals can be observed.

A. Brief Outline of USADR system.

FM-1 NOMAC System Assumptions

The digital communications signaling system for the basic FM-1 system presented in [13] is assumed to be a 48 channel noise modulation (NOMAC) signalling scheme, wherein each channel is binary with biorthogonal coding. Each signalling waveform is assumed to be 192 samples from a random number generator, subsequently band pass filtered to control the baseband spectrum. The details of the filter are in a patent application, as well as how the filtered waveforms are then orthogonalized to remove filter induced correlation.

In order to avoid these unknowns, our baseband computer simulation operates without the bandpass filtering so as to maintain maximum signal orthogonality. We could of course design a BPF to match the plots supplied in [14], however we would have no way to the orthogonalize the resultant 48 band pass filtered signalling waveform suite. In any event this conservative approach bounds the performance, which is already degraded from the optimum for reasons later discussed in this report.

If the bit rate per channel is 8 kbps the aggregate channel bit rate is 384 kbps. At these rates the random number sequence would require a sampling rate of 1.536 MHz. If a rate 1/2 Forward Error Correction (FEC) algorithm is utilized to guard against channel impairments, then a 192 kbps information data rate is achieved.

The simulated results are compared to mathematical theoretical results available for M-ary biorthogonal signalling to quantify the performance loss. We use theoretical canonical performance bounds so that we are insulated from any specific hardware implementation details, which are not provided. Additionally, use of theoretical m-ary bounds provides an absolute limit on the best performance that can be achieved; any real world system or simulated system can not exceed the theoretical bound, although systems without cost constraint may well be designed to operate near or at theoretical limits. The signal spectrum is not estimated as we can not accurately model the baseband filtering applied to the pseudo random number sequence used as the signalling suite. It may be noted, however, that clearly the noise waveforms can be filtered to meet a desired spectral mask, but the performance will degrade measurably unless the filtered signal suite is perfectly orthogonalized. Additionally, a 48 signal overlay suffers from the impairments common to any CDMA system in that a performance floor is set by the processing gain and the number of users, even if infinite signal to noise ratio (SNR) is assumed.

The receiver for NOMAC digital systems is straightforward, requiring 48 correlators, each operating on 192 waveform samples, an operation that can be achieved in real time DSP using FFT based high speed correlation algorithms, programmed into commercially available DSP chips.

FM-1 Enhanced System Using Gold Codes

The FM-1 enhanced system, based on a suite of 32 channel, 64 bit (assumed) Gold codes, and 32-ary signally, was simulated to estimate the spectral content of the transmitted signal. The processing gain of this spread spectrum system is directly proportional to the length of the Gold codes used; clearly the higher the processing gain the better multipath tolerance and narrow band interference rejection, but also the high chip rate that must be supported by the channel. The processing gain of any spread spectrum system, including this spread spectrum system, has absolutely no effect on performance in the presence of additive white Gaussian noise; this means that any interference from adjacent channel IBOC systems that appears as white Gaussian noise or statistically similar is not mitigated by increasing or decreasing processing gain. Processing gain for direct sequence spread spectrum (DSSS) systems is beneficial only to combating multipath and narrow band interference. The simulation is used to estimate performance degradation to the self interference effects of a 32 signal overlay without benefit of significant spectrum spreading and compared to theoretical bounds as a cross check. Specific hardware implementations will cause additional unavoidable losses in processing gain of as much as 3 dB.

As determined from [13], we assume availability of a 96 kbps stereo audio codec algorithm, a rate 1/2 FEC, and a bandwidth efficiency of at least 2 bps/Hz, so that the signal will fit the 100 kHz spectral mask set by the FCC. Therefore, each of the 32 channels must operate at a bit rate of 6 kbps to produce an aggregate channel bit rate of 192 kbps. We can constrain the baseband spectrum to 100 kHz by use of raised cosine pulse shape filtering (or any similar method such as Gaussian filtering in GMSK), which in turn constrains the per channel chip rate. For narrow band interference, self interference protection, and multipath robustness we desire the Gold code length to be as long as possible to increase the spread spectrum processing gain. The self interference protection is important because 32 signalling suites are transmitted together, thus the receiver sees the mathematical equivalent of a 32 user CDMA system with perfect power control. The length of the spreading code does not alter BER performance against Additive White Gaussian Noise (AWGN). If we set the Gold code length to 64 bits, then use of 32-ary biorthogonal signalling induces a chip rate of 76.8 kcps, a rate that with proper pulse filtering can operate within a 100 kHz bandwidth (baseband). The processing gain is 10Log(64/5) = 10Log(76.8 kcps/6 kbps) = 10Log(12.8) = 11 dB.

Compared to the FM-1 NOMAC system the complexity is significantly increased. The receiver requires (32)(16) = 512 (assuming biorthogonal signalling) 64 bit correlators in order to decode the channel bits, then a rate 1/2 decoder to decode data bits. By contrast, FM-1 NOMAC required a set of 48 correlators, each 192 samples in length.

Biorthogonal signalling requires coherent demodulation as the correlators much detect the sign of the correlation pulse; if orthogonal signalling is used a noncoherent demodulator may be used, but the number of correlators doubles to 1024, assuming the system we have constructed.

B. Input Data

1. a) FM-1 NOMAC System

A pseudo random sequence length of 192 is assumed (not stated in USADR report) as the example autocorrelation plot was 384 samples in length since a discrete time autocorrelation estimate typically doubles the length of the discrete random data. The pseudo random sequence sample rate was stated in the USADR report as 1.536 MHz; assuming a 192 pseudo random number sequence produces a per channel rate of 8 kbps. If the aggregate channel rate is 384 kbps as stated in USADR report, then if each channel operates at 8 kbps it is required to overlay 48 channels. A pseudo random number sequence enjoys infinite bandwidth, thus baseband filtering is required to meet FCC spectral mask requirements; for obvious reasons the filtered pseudo random sequences must be re-orthogonalized since filtering a random sequence induces correlation between signal samples. The filtering is mentioned, but not specified, in the USADR report. The M-ary communication system theoretical performance assumes perfectly orthogonal signals, thus serves as a bound on achievable performance. Any realistic system can approach, but never exceed these bounds; such an analytical approach is independent of specific system implementation and bounds performance.

1. b) FM-1 Enhanced Gold Code System

The example autocorrelation and cross correlation plots in the USADR report indicated use of 127 bit Gold codes. The chip rate (ie, the rate the Gold code bits are transmitted) was stated in the USADR report to be 75.6 kcps. It was further stated that 32-ary system could be constructed with a 32 channel overlay to meet the channel rate requirements stated to be 192 kbps.

If we assume a 32 channel system, with an aggregate channel rate of 192 kbps, then each channel must operate at 6 kbps. If we set the Gold code length to 63 bits, vice 127 bits, and use the stated 32-ary signalling, then the chip rate is (63/5)(6 kbps)=75.6 kcps, which matches the chip rate stated in the USADR report. A system constructed in this manner would exhibit a progressing gain of approximately 11 dB.

2. Parameters that are needed but not supplied by USADR.

a) Parameters Required To Analyze Digital Communication Systems

Since modern day digital communications systems usually require power efficiency and bandwidth efficiency as a primary design constraint, coherent digital modulation techniques are used extensively. Common coherent digital modulation signaling methods are binary phase shift keying (BPSK), quadrature phase shift keying (QPSK), minimum shift keying (MSK), Gaussian MSK (GMSK), M-ary PSK, M-ary Quadrature Amplitude Modulation (QAM), and orthogonal/biorthogonal signally (may be binary or M-ary). The modulation methods (BPSK, QPSK, MSK, and orthogonal/biorthogonal) provide the nearly the same BER performance, optimal against an additive white Gaussian noise (AWGN) channel without multipath, but differ in complexity and power spectra characteristics.

The bandwidth of a digitally modulated signal and the shape of the signal spectrum are determined by the type of digital modulation, type of baseband data encoding utilized, and the data rate. The data itself is commonly encoded as non-return to zero (NRZ) or Manchester (biphase) format, although other data encodings are possible. Additionally, it is a near certainty that the digital data will be subjected to source coding in order to reduce the baseband data rate. For example, digital broadcast audio data will generally be coded to a lower bit rate via routine parametric coding algorithms such as Code excited Linear Prediction (CELP), prior to modulation. It is common to combine source coding with some form of channel coding for error protection (the source coded bits with less redundancy are more sensitive to bit errors than digitized raw audio). A multipath communications channel is considered very harsh; thus, in order to ensure reliable communications, channel coding may be invoked to protect some or all of the source encoded bits. Although source coding decreases the data rate, channel coding increases the data rate, so the two coding strategies are at odds with each other. Both are usually necessary, however, so a digital system receiver designer must deal with the added complexity involved from a system design viewpoint. As might be expected, the higher complexity modulation types such as MSK offer greater bandwidth efficiency, thus allowing higher data rates to be accommodated for a given RF bandwidth. The trade-offs between data rate (a function of baseband data rate, source coding, and channel coding), modulation type, and processing complexity are virtually endless.

b) Parameter Assumptions - Line Coding

A line code maps the logical binary data to analog voltage levels for subsequent modulation onto a carrier and transmission over the ether. Common lines codes are Manchester, NRZ, and RZ, each possessing advantages and disadvantages. NRZ is the most bandwidth efficient and enjoys the lowest BER, but requires DC coupling, and performs the poorest within the bit synchronization algorithm in the digital receiver. Manchester coding possesses the same BER as NRZ, is easy to bit synchronize, and admits AC coupling, but requires double the bandwidth of NRZ. RZ is a compromise between NRZ and Manchester, trading off BER performance for AC coupling and bandwidth requirement somewhere between NRZ and Manchester.

The original NOMAC system does not use a line code as its a binary valued noise modulation system. The enhanced system was assumed to use NRZ line coding in order to minimize bandwidth.

c) Baseband Pulse Shaping

The originally proposed NOMAC system does not utilize binary pulses, rather uses pseudo random noise sequences as a signal suite. These sequences (infinite bandwidth) must be filtered to control the spectral content. The baseband filter parameters are unknown; in any event, if we constructed a filter we then could not re-orthogonalize the filtered sequences. As a consequence the filtering was not used so that our simulation operates in a more ideal manner than a real system and serves as a performance bound. Additionally, we can theoretically bound performance from M-ary detection theory, and compare our simulation to a mathematically perfect system.

The spectrum of rectangular pulses is infinite in extent with a strong sin(x)/x sidelobe structure, not suitable for any digital system that is required to operate in a limited spectral mask. Common pulse shaping filters to limit spectral occupancy and/or to mitigate intersymbol interference are Nyquist filtering, Gaussian filtering, and raised cosine filtering, although many other filters may be used. We chose arbitrarily a square root raised cosine baseband pulse filter as it is specified in the IS-54 North American Digital Cellular Standard. Such a filter may be designed to constrain the spectrum as tightly as desired, however, the receiver performance degrades as the bandwidth is decreased, as might be expected. We used a compromise value of beta-1.0 filter, which was the widest bandwidth that could just meet the spectral mask easily.

d) Data Rates

The digital data rate simply refers to the number of bits per second transmitted across the channel. The higher the data rate the wider the bandwidth all other factors being equal. We chose data rates to match data rates stated in the USADR report as previously discussed in this report.

e) Digital Modulator

Our simulation operates at baseband (view the bandpass spectrum as frequency shifted to baseband) as do all digital communication system simulations as it is not feasible to sample RF waveforms accurately. Imagine sampling an 88 MHz FM carrier at the Nyquist rate of 176 megasamples per second and trying to run the computer program. Baseband analysis is equivalent to bandpass RF analysis. We note, however, that the USADR system, assuming biorthogonal signalling, must implement a coherent modulator/demodulator. Furthermore, the baseband spectrum must translate (modulate) to the FM radio band without spectrum doubling so that DSB can not be used (unless chip rate is cut in half).

f) IF/RF Filtering

A digital communications transmitter requires IF and RF filtering to control spectral content. We possess no knowledge of the filtering applied. The simulations operate at baseband and do not include additional filtering, thus we provide a performance bound.

APPENDIX C. The model

The model simulation was coded in Matlab, a commonly available software package based on linear algebraic mathematical functions. Features of Matlab appropriate to digital
communication system modeling include random number generators, correlators, filter design functions, and FFT functions to estimate frequency content of random waveforms.

This simulation (and all digital communication simulations) operate at baseband so that the simulation sampling rates can be managed. Sampling at RF rates is simply not possible or even desirable.

All system parameter assumptions have been previously stated or are stated in the following sections. The model can be used to estimate baseband frequency content, and the effects on BER due to signal overlay or the effect on BER due additive white Gaussian noise specified as energy per bit / noise power (Eb/No). Eb/No is the SNR metric of choice as it is parameter easily related to specific system parameters such as data rate and carrier power. The additive white Gaussian noise disturbance is appropriate to modeling receiver thermal noise, and the noise induced by a spread spectrum adjacent channel IBOC signal.

The model is not currently coded to predict performance degradation due to multipath interference; such a channel model is comparatively complex and could be added with additional contract funding at a later date.

Specific baseband pulse filtering can be easily coded; the only filter available at this time is a square root raised cosine filter. Specific Gold codes used in a proposed system could be easily added if supplied; exact Gold codes used would be highly desirably if detailed performance results were desired in the future.

We verify (specific examples in other sections of this report where appropriate) the simulation code by comparing BER to Eb/No from the simulation to known perfect theoretical results from M-ary detection theory. The theory provides an absolute bound that can not be exceeded. The simulation will always be slightly degraded but close to theoretical results as the simulated signals will not be perfectly orthogonal and are finite in length. Any specific hardware simulation can be made to operate close to simulated results, although other losses may be incurred such as correlator insertion losses. For example, a proprietary system MDS possesses knowledge of was designed for a
15 dB processing gain, but measured about 11 dB in the hardware laboratory.

Comparison of simulation results to theoretical results is very powerful. No system can exceed theory; if the simulated system then operates at the edge of say some spectral mask, then a prototype system can do no better. This allows very broad statements to be made with confidence about any enhancements offered.

1. FM-1 NOMAC IBOC

In Figure 1. a comparison is shown for the theoretical biorthogonal (binary case) BER predictions (solid line), and the results from our baseband computer simulations (X's). We note a several dB loss of performance, which is expected as signal orthogonality is not perfect with finite length pseudo random sequences; the theoretical BER curve assumes infinite length signalling waveforms. Figure 1. is instructive and serves to validate our baseband computer simulation for NOMAC communication systems.

Several sources of performance degradation may be noted from our canonical transmitter/receiver system. First, the processing gain is 10 log10 (192 / 1) = approximately 23 dB, therefore the system will suffer from narrow band interference in its passband that exceeds about 23 dB, increasing the bit errors. Second, it requires an infinite length signal suite in order to achieve the perfectly uncorrelated white noise characteristic that the mathematical analysis assumes. A signal 192 samples in length, even if generated by a very good random number generator, possesses significant cross correlation structure, increasing the bit errors. Third, the waveform synchronization is critical; any mismatch between the signalling waveforms in time against the receiver stored waveforms in the correlator bank will degrade the error performance from theoretical predictions. Fourth, the idea of overlaying 48 noise waveforms is mathematically equivalent to a 48 user CDMA system with perfect power control. We can bound the performance using established CDMA multi-user access equations, then assess the degradation due to loss of orthogonality over a finite length noise signal via our computer simulation (the simulation includes the entire signalling suite of 48 waveforms. The text of [13] refers to a patent for perfectly orthogonalizing all 48 band pass filtered waveforms, however, that would only be true for perfect synchronization in time. Any synchronization errors would induce correlation amongst the signalling waveforms.

A more significant loss of performance is seen when 48 channels are summed in order to increase the aggregate channel data rate for transmission of stereo audio. In Figure 2., which plots the simulated single channel BER data (X's) against the simulated 48 channel BER (O's), we see that nearly 15 dB additional Eb/N0 is required for the same performance.

2. Enhanced IBOC System

We note that the enhanced system will suffer BER performance degradation as well, although not to the extent of a NOMAC system. Gold codes are nearly orthogonal, thus can approach theoretical limits of signal overlay. A more detailed multi-user analysis (ie, co-channel interference performance) is presented in the next section. We did, however, use an asymptotic mathematical expression, based on perfectly orthogonal codes, with noise power set to zero (equations in next section) to predict that the probability of bit error should be approximately .04 with 32 users, each with equal power. A long simulation run (32 users, nearly orthogonal random binary 64 bit codes, no added noise) estimated the Pe to be .07. This is in very close agreement, the difference due primarily to statistical variation and not quite orthogonal codes; such a simulation test serves to validate our FM-1 enhanced baseband simulation, providing extra confidence for use in spectral estimation, a more important issue in IBOC system analysis.

It is very important to note that perfect code synchronization is available in the baseband simulation to produce the results in the preceding paragraph. Indeed, through varying the simulation parameters, it was found as expected that performance decreased rapidly as increasing synchronization error (as a fraction of a chip) was inserted. This effect is easily seen from inspection of the autocorrelation function of a Gold code [13].

Based on Pe equations readily available [12], for M-ary biorthogonal signalling, the theoretical BER can be predicted and used as an upper bound on system performance. The Pe expressions for M-ary biorthogonal signalling are not closed form, awkward to evaluate, so we have reprinted a plot (Figure 6.) from [12] to summarize expected probabilities of bit error. These curves assume perfectly orthogonal signals, perfect synchronization, an AWGN channel, and rectangular baseband pulse shapes, and ideal matched filter receiver architecture. Obviously, real world performance will degrade in the presence of a mobile propagation channel, pulse shaping, and synchronization errors.

The performance loss in a real world fading mobile channel is significant. To present a simple comparison, we reprinted known theoretical results from [11] in Figure 7. Comparatively few theoretical results are available for fading channels, thus simulations are used to quantify performance predictions for specific modulation schemes, different signalling schemes, different baseband filtering methods, and varying mobile channel conditions, as all these factors affect the performance in highly nonlinear ways.

3. Baseband Spectrum Estimates

We used computer simulations at baseband in order to estimate the spectrum of the Gold code based enhanced FM-1 system. The NOMAC system was not analyzed for spectral content as its spectrum is controlled purely by the bandpass filtering (unknown filter parameters) of the random noise signalling waveforms, thus classical baseband pulse shaping can not be imposed.

The enhanced system that makes use of Gold codes represents a straightforward modification to our canonical simulation as it's simply a signalling waveform change; approximate Gold codes have been added as appropriate and simulation results used to estimate the baseband signalling spectrum, assuming raised cosine pulse shaping for spectrum control. Additionally, Monte Carlo simulations were performed to estimate performance degradation due to signal overlay self interference, with results consistent with asymptotic theoretical predictions.

The spectrum of a digitally modulated signal is governed by the data rate, baseband pulse shaping, the modulation scheme selected, and additional RF filtering applied in the transmitter (refer to Appendix for additional detail).

Any IBOC system will benefit greatly from lowering the data rate to rates below 100 kbps, obviously, and the available codecs are moving in that direction. Here, we assume a 96 kbps codec will be available for use in the near future. High quality compression of wideband audio (typically 7 kHz or 20 kHz bandwidth) will become increasingly important for digital commercial radio broadcasting and Integrated Services Digital Network (ISDN) applications.

128 kbps 20 kHz stereo is now available, and it APPARENTLY sounds quite reasonable, though not identical to a CD original. One product that offers this is the Comrex DX200. It is not known exactly how their codec works, but the audio coding approach that seems to be most popular these days might be called "suband-transform" coding. The transform would be something akin to a DCT, properly sized and overlapped to get good transient response, and the suband nature allows differential bit-allocation across the band in an attempt to place the quantization noise where it is least audible. It is our belief that MPEG now has standardized 3 layers of audio coding, and is working on a fourth. The sound system for digital television goes by the name of Dolby AC-3.

The bandwidth available to an IBOC system in a single sideband is 100 kHz; if a data rate of 192 kbps is desired (assume rate 1/2 FEC, and a date rate of 96 kbps), then the spectral efficiency is set at 2 bps/Hz. Digital modulators such as 16-ary PSK and 16-ary QAM can achieve 2 bps/Hz, and 32-ary PSK/QAM can achieve 2.5 bps/Hz. The penalty paid for M-ary PSK style system to achieve high bandwidth efficiency is reduced power efficiency. QAM is more power efficient than PSK, however, receiver complexity is increased. The FM-1 enhanced system, however, proposes to use M-ary biorthogonal signalling, with signal overlay, as discussed next.

For the FM-1 enhanced IBOC system, however, wherein biorthogonal signalling is used, vice PSK signalling, the bandwidth efficiency is achieved by (apparently) careful baseband pulse shaping, and signal overlay of 32 channels, each channel chip rate being 76.8 kcps (assumed) and a data rate per channel of 6 kbps (assumed). The signal overlay does not affect the power spectral density of the baseband signalling suite. M-ary orthogonal signalling or M-ary biorthogonal signalling increases power efficiency as M increases, but at the expense of bandwidth efficiency, thus the signal overlay procedure to recover "lost" bandwidth. We will find later in this report that baseband pulse shape filtering is critical to meeting spectral mask requirements, so that interference to adjacent channel IBOC signals and FM host main channel signals is minimized. By comparing the theoretical bandwidth efficiency of M-ary biorthogonal signalling to system bit rate requirements we will find the theoretical limit is exceeded, mandating use of a pulse shaping filter.

From Figure 3. we see that use of a beta=1 raised cosine filter [11] the baseband spectrum at 100 kHz is 50 dB down from the passband level. The parameter beta, or filter rolloff factor can be varied to achieve spectrum control, by trading off correlator losses and bit synchronization complexity if the bit stream is tightly filtered. For example, the IS-54 North American Digital Standard specifies a rolloff factor of .35 for a raised cosine pulse shaping filter.

3. Co-Channel Interference Analysis

In a wireless spread spectrum Code Division Multiple Access (CDMA) environment signals from a number of users on the same frequency channel arrive at the receiver input. A correlation receiver with a pseudo noise (PN) code matched to some desired user is used to separate one signal from the other signals; since the other users are assigned PN codes approximately orthogonal to all others the net effect is a background nearly white noise level. For this IBOC scenario of the NOMAC FM-1 system, the co-channel interference is actually self interference from the 48 channel signalling suite. In the FM-1 enhanced system based upon Gold codes the self interference consists of 32 channels of signalling. If the number of co-channel users is large enough so that the Central Limit Theorem may be invoked an expression may be obtained for BER that is very simple, especially if perfect power control is assumed (mutli-channel signalling as in this IBOC analysis is equivalent to perfect power control), wherein all signals arrive at the receive with equal power. For this report we will assume Gaussian statistics (number of users "large", say five or more at least), and consider both the equal power scenario and the unequal power scenario.

A widely accepted expression for Probability of Bit Error (Pe), that has held up in field trials is

Pe = Q{[3N/(P1/P0 + P2/P0 + ... +PK-1/P0 + N0/2TbP0)]1/2}

where Q(.) is the standard "Q-Function," N is the length of the PN code, Tb is the message bit interval, N0 is the noise power in Watts/Hz, P0 is the power in Watts of the desired signal, and Pk, k = 1, 2, 3, ... , K-1, is the power in the undesired signals representing the co-channel interference power. This expression assumes the interfering signals are of fixed number, K-1, and constant but unequal power. In a wireless scenario that is co-channel interference limited rather than noise limited, N0 can be set to zero for convenience; this is useful for producing quick estimates of performance degradation due solely to co-channel interference.

A very simple expression obtains by setting noise power to zero and assuming all signals arrive at the receiver with equal power Pk = P0 Watts. This case results in

$$Pe = Q[(3N/(K-1))1/2]$$

from which it is clear that either a 48 channel or 32 channel signal overlay will experience self interference. For example, for $K = 48$ signals, $N = 192$ PN code length, we find Pe is approximately .0006; adding noise, N0, to this example would decrease the value of the argument in Q(.), increasing Pe measurably, as would be expected. The enhanced FM-1 system, consisting of 32 signals without additional spectrum spreading would be expected to experience a Pe = .04 using a 32 bit code for computational savings. Our baseband simulation estimated the BER at approximately .07, so we are quite confident in our simulation accuracy. We can easily insert any desired code length in the baseband simulation or code construction desired.

5. Adjacent Channel and Host Interference Issues

The IBOC signal must be positioned in the FM sidebands beneath a spectral mask as determined by the FCC so as to not interfere with either the host FM signal or adjacent channel signals. The appropriate spectral mask will not be repeated here as it is commonly available in other reports (47 CFR 73.317, Appendix B), but basically requires the digital sideband to reside within 120 kHz to 240 kHz from the main channel (host) unmodulated carrier with spectral magnitude at least 25 dB below the host carrier. Note that this mask refers to spurious emissions; digital sideband spectra that meet the minimum criteria of this specification may interference with adjacent channel IBOC stations and/or host FM main channel audio, thus a tighter bandwidth may well be required in practice for an IBOC signal. A 32 channel, 32-ary 192 kbps digitally modulated signal can not be viewed as a spurious emission.

Other potential problems resulting from the digital sideband signal include interfering with a third harmonic of the 38 kHz subcarrier (receiver manufacturer specific) and interfering with a potential 92 kHz SCA.

The enhanced FM-1 IBOC spectrum is governed by the baseband pulse filtering, the per channel bit rate, the 32-ary biorthogonal modulation choice, and any subsequent RF filtering. The power (area beneath the PSD) is obviously a function of the bandwidth, spectral shape, and peak transmitted power in the digital sideband.

As can be seen from Figure 3., the baseband signalling spectrum can be tightly controlled by careful pulse filtering, and selection of bit rates. The spectrum measured by and reported in [14] indicates clearly that spectral energy is visible below the 120 kHz lower limit at some level difficult to accurately gauge, and exceeds the upper 240 kHz limit by some margin, even exhibiting energy beyond 250 kHz. This spectral energy would adversely affect an adjacent channel IBOC signal. It is not known by this author, with respect to the spectral plot viewed from [14], what baseband pulse filtering was used (if any), what precise bit rates were used, and what RF filtering was imposed. The spectrum can be modified significantly by varying such design parameters; the baseband spectrum estimate shown in Figure 3., indicates that use of a raised cosine pulse filter, combined with a per channel chip rate of 75.6 kcps (assuming 32-ary biorthogonal modulation and 64 bit Gold codes), may bring the digital sideband spectrum within the mask boundaries.

The baseband pulse filtering can be critical to spectrum control. In our simulation, we created a 32 signal overlay of 64 bit codes, both filtered (beta = 1 raised cosine filtering), and unfiltered. The spectrum shown in Figure 4. is unfiltered, and the excessive spectral energy outside the mask limits is clearly visible. By comparison, the spectrum shown in Figure 5., estimated from a 32 signal overlay of raised cosine filtered 64 codes is very well behaved, falling well with the 120 kHz to 240 kHz mask limits, with greater 50 dB attenuation from passband at the mask boundaries. As a reminder these spectrum were produced by a baseband simulation with 76.8 kcps per channel chip rate. The baseband signalling suite is subsequently modulated to the sidebands, thus close attention must be paid to any potential bandwidth expansion from the modulator. We assume the baseband spectra is translated to the sidebands used for IBOC transmission without bandwidth expansion by use of DSB modulation. This modulation scheme would replicate the digital signal redundantly into a lower and upper sideband, allowing use of the diversity idea proposed in [13] in order to improve digital audio performance.

Indeed, the theoretical limit (assuming no pulse filtering) is somewhat exceeded by the M-ary biorthogonal signalling system proposed, thus pulse filtering is absolutely required to meet spectral mask constraints. From [12], the theoretical maximum bandwidth efficiency for M-ary biorthogonal signalling using rectangular pulses is 4Log2M/M bps/Hz, which for 32-ary signalling is 5/8 bps/Hz. A 32 signal overlay plus spectrum spreading of 64/5 results in a maximum bandwidth efficiency of (5/8)(32)(5/64) = 25/16 bps/Hz, slightly less than the 2 bps/Hz required for this proposed IBOC scheme to fit safely within the spectral mask, thus not causing interference to adjacent channel IBOC systems or the host FM station main channel. Note that 25/16 bps/Hz is a theoretical maximum bandwidth efficiency for 32-ary biorthogonal signalling, and the real world efficiency will be lower. From the preceding analysis we see that baseband pulse shaping is mandatory.

The power in the digital sideband(s) can be set to bring the spectrum beneath the mask upper limit and/or control interference to subcarriers to limit the PLL decoder problems. Obviously, lowering the power lowers the per bit Eb/N0, with a resultant increase in BER.

The IBOC FM-1 enhanced system design as postulated in this report, based upon bit rates and signalling specifics detailed in the system description section of this report, would exhibit a different digital sideband spectrum than apparent in [14]. For example the pulse shaping filter parameter(s) can be varied to control baseband spectra, the RF filtering can be used to control transmitted signal spectra, the FEC rate can be used to lower the aggregate channel rate, the length of the Gold code can be used to control the per channel chip rate to advantage, and the M-ary signalling can be used to BER against baseband bandwidth. These design factors are highly interrelated of course, and any measure used to reduce digital sideband bandwidth will inevitably increase the BER, potentially below accepted performance expectations. Ultimately, however, the promise of improved codecs (ie, lower bit rate), will buy performance trades in favor of any IBOC system in use.

6. Narrow Band Interference Analysis

One expects that the IBOC receiver would not experience significant interference from NB signals, rather most interference should result from spectral leakage from the host FM station, and adjacent channel FM stations with IBOC capability. In the event NB interference may be an issue, a brief summary of the appropriate equations is given.

The output SNR in dB, (SNR)o, for a generic correlation DS receiver, may be related to Processing Gain (PG) in dB, and input SNR in dB, (SNR)i, by the expression

$$(SNR)i = (SNR)o - PG$$

where Processing Gain is defined as the ratio of spread spectrum signal bandwidth to message signal bandwidth. If we assume rectangular pulses for message bits, I/Q modulation, for the basic FM-1 NOMAC system, the processing gain in dB is given by

$$PG = 10 \, Log(192) = 23 \text{ dB approximately,}$$

although the true PG is usually lower than the theoretical due to a residual carrier component from the modulator. The enhanced FM-1 system, based upon a set of 32 waveforms consisting of 32-ary 64 bit (assumed) coding, is a spread spectrum system with a low processing gain of 10Log(76.8/6) = 11 dB, thus NB interference rejection is limited to 11 dB.

The output SNR of a correlator receiver is given by

$$(SNR)o = P/[(N0/2Tb) + (J/PG)]$$

where P is the power in Watts at the receiver input due to the desired transmitted DS signal, N0 is the noise power in Watts/Hz, Tb is the message bit interval, PG is the processing as previously defined, and J is the power in Watts due to a NB interference source.

Substituting Eb = PTb, and rearranging terms slightly, we can write

$$Eb/N0 = [1/2 + J/(N0B)](SNR)o$$

where B is the total DS signal bandwidth in Hz, in order to directly relate BER to NB interference power in Watts and noise power in the receiver bandwidth. We see from this expression that if the interference power, J, is narrow band then increasing the spreading (ie, increasing B), improves performance. Note, however, that if the interference power is wide band then increasing the spreading does not improve performance because the interference power increases with the bandwidth.

7. Digital Receiver Complexity

Functionally, the receiver must accomplish several tasks generic to most digital data communication systems, although some processing tasks may not always be required. The receiver may be required to extract and track doppler information, and must always demodulate the carrier (at IF) in order to produce a soft bit stream suitable for further processing; obviously, the soft bit stream will require further processing before hard bit decision are declared. For a digital signal processing based receiver, the digital data demodulation may be performed in software or dedicated digital hardware.

The retrieval of data symbols from the received signal will involve demodulation of data modulated via binary phase shift keying (BPSK), quadrature phase shift keying (QPSK), offset QPSK (OQPSK), minimum shift keying (MSK), continuous phase modulation (CPM), Gaussian minimum shift keying (GMSK), frequency shift keying (FSK), quadrature amplitude modulation (QAM), amplitude shift keying (ASK), or a wide variety of other related modulation schemes in use today for modern digital communications systems [1,10,11,12]. As bandwidth efficiency, and power efficiency, often competing design constraints, are of paramount importance for digital data communication systems, ever increasing digital signal processing based receiver complexity is a certainty.

Data demodulation is a necessary step in extracting digital data from the received signal, but many other processing steps are usually invoked to improve receiver performance, especially in harsh multipath propagation communications environments. For example, so called RAKE receivers [9] take advantage of the information contained in multipath components to improve bit error rate performance at the price of processing complexity. Almost all time division multiple access (TDMA) systems require channel equalization to mitigate the effects of intersymbol interference [8]. Code division multiple access (CDMA) systems may not require sophisticated channel equalization, but do require code correlation processing. Literally hundreds of research papers and technical books are available on these advanced topics. Since the multipath communications channel is a time-varying system, adaptive digital filter processing is required for best performance, further increasing the complexity of modern digital data communication systems. If additionally, the receiver is based upon digital signal processing methods, the complexity is increased more so as the algorithms must adapt and execute in real-time.

A significant complexity issue for the USADR proposed enhanced system is the requirement for 512 correlators, each 32/64/128 bits in length, depending on the specific design. Correlators, perhaps a SAW device, are comparatively expensive with significant power consumption and insertion losses . In this case, 512 correlators is an impractically high number of separate correlators, so these would have to be implemented in some sort of very high performance digital processing system.

7. RAKE Receiver Complexity

In spread spectrum CDMA systems, due to the relatively high bandwidth of the signals (of course with increased complexity) several different propagation paths can be resolved. In such cases the RAKE receiver [9] can be put to good use in order to decrease the bit error rates. The RAKE receiver in its simplest form is a linear weighted sum with time-varying coefficients that are a function of the attenuation and delay of the individual multipath components. Since multipath propagation is highly time dependent, especially in a mobile communications environment, the coefficients must be estimated from an estimated channel impulse response and updated in real-time to accurately model the current channel propagation characteristics.

The coefficients are a function of a linear convolution that can be implemented via standard FFT based high speed convolution algorithms with well defined computation complexity. The complexity trade-offs will be a function of data block size (ie, FFT length) and how often the coefficients require update, which is a function of channel dynamics; computer simulation must be utilized to model the specific multipath communications scenario and define data block sizes and update rates. The preceding operations lend themselves to conventional ASIC implementation that can handle approximately 100 million multiply/add operations per second.

A specific example of a RAKE receiver DSP design is given in [3], for a spread spectrum communication systems application in a multipath environment. A comparatively high chip rate is used (approximately 20 Mchips/second) in order to resolve the relatively small delay spread of the indoor channel in the specific application considered. For this application a data rate of 16 kb/s is invoked, the digital modulation chosen is BPSK, and Gold codes are used for spreading, a common choice; the receiver is capable of resolving 8 multipath propagation paths, and the correlator operates digitally at baseband in time in parallel over each of the 8 RAKE 'arms.' The receiver is based upon two time integrating correlators providing 8 parallel channels into a TMS320C25 executing with a 100 nsec cycle time.

Figure Titles

Figure 1. Simulated BER For FM-1 NOMAC System Vs. Theoretical BER For Binary Biorthogonal Signalling.

Figure 2. Simulated BER For FM-1 NOMAC Single Channel Performance Vs. 48 Channel Signal Overlay.

Figure 3. Baseband Spectrum Of NRZ Gold Code Sequence With Beta = 1.0 Raised Cosine Filtering.

Figure 4. Baseband Spectrum Of NRZ Gold Code Sequence With Overlay Of 32 Signals, No Filtering.

Figure 5. Baseband Spectrum Of NRZ Gold Code Sequence With 32 Signal Overlay, Beta = 1.0 Raised Cosine Filtering.

Figure 6. Theoretical BER Curves For M-ary Biorthogonal Signalling (Reprinted From [12]).

Figure 7. Theoretical BER Curves Comparing Performance In AWGN Vs. A Fading Channel (Reprinted from [11]).

APPENDIX REFERENCES

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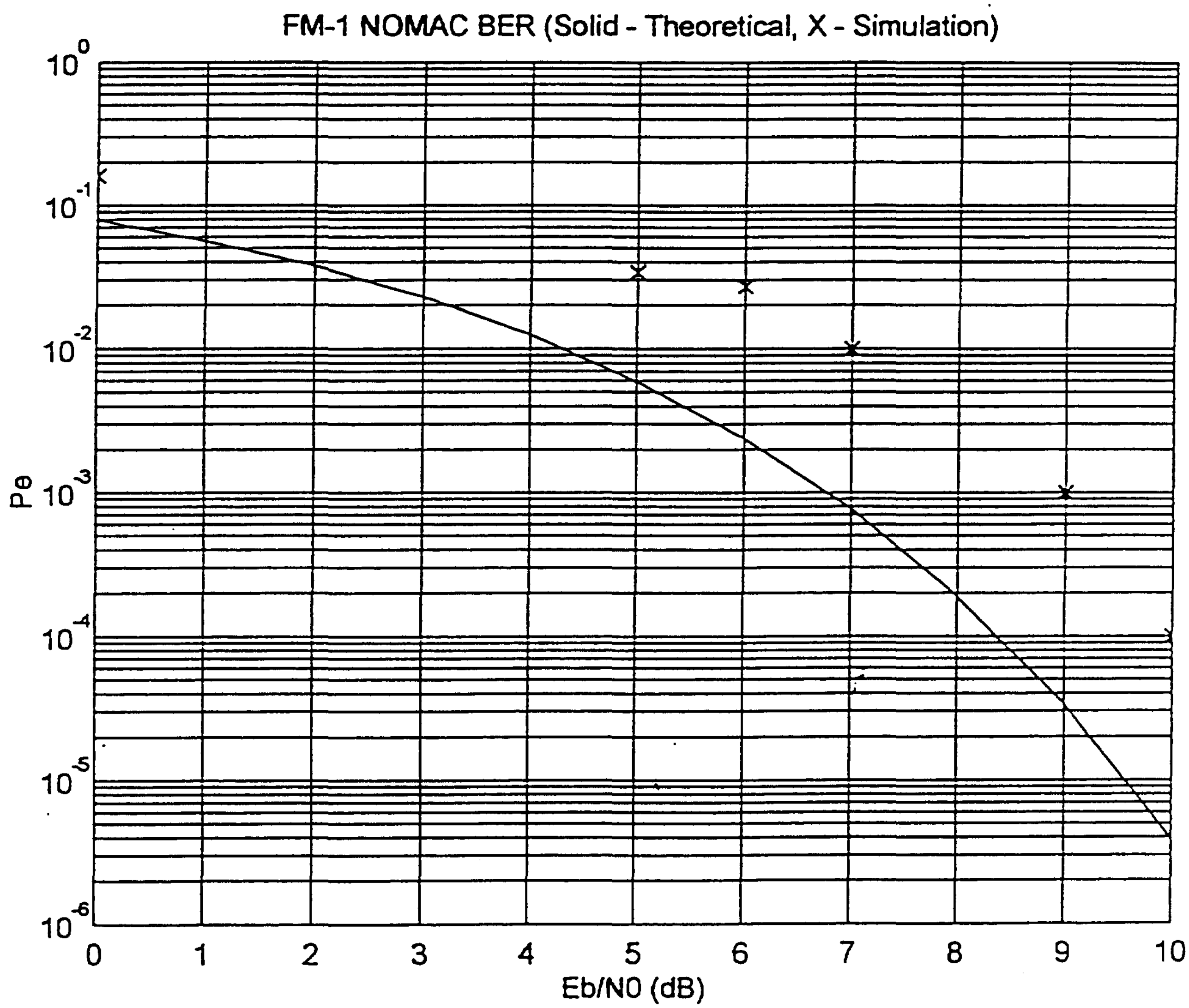

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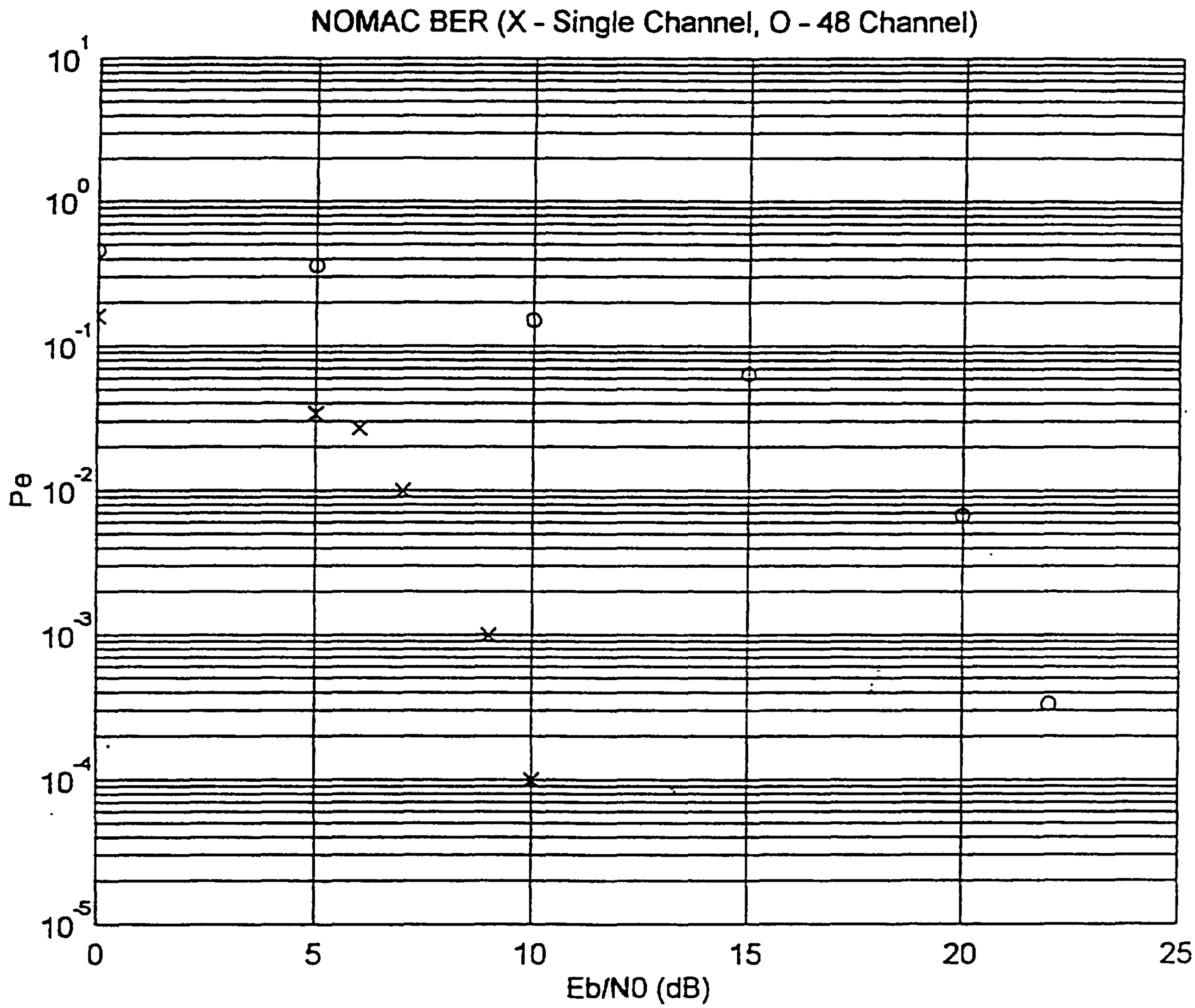

Figure 2. Simulated BER For FM-1 NOMAC Single Channel Performance Vs. 48 Channel Signal Overlay.

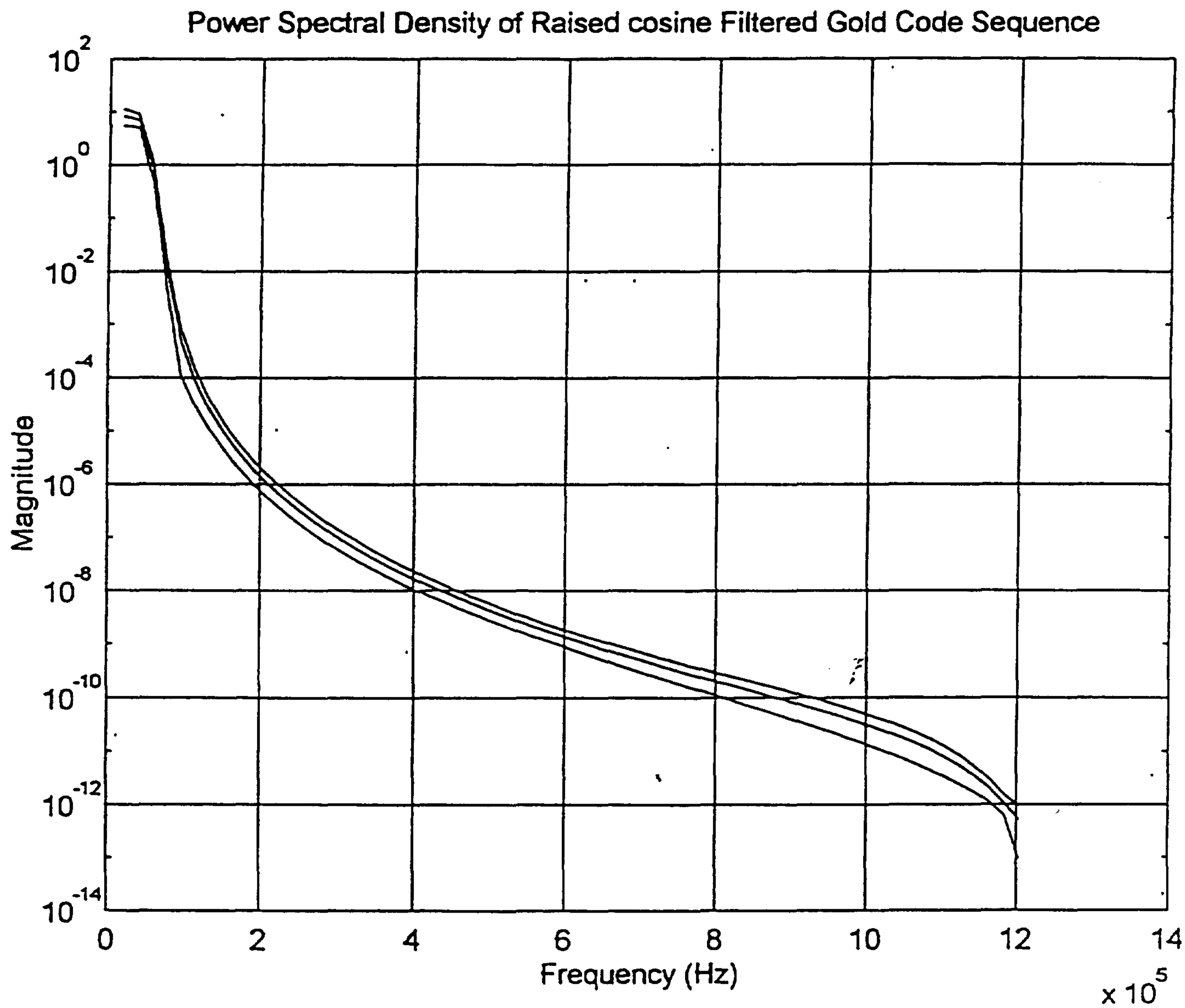

Figure 3. Baseband Spectrum Of NRZ Gold Code Sequence With Beta = 1.0 Raised Cosine Filtering.

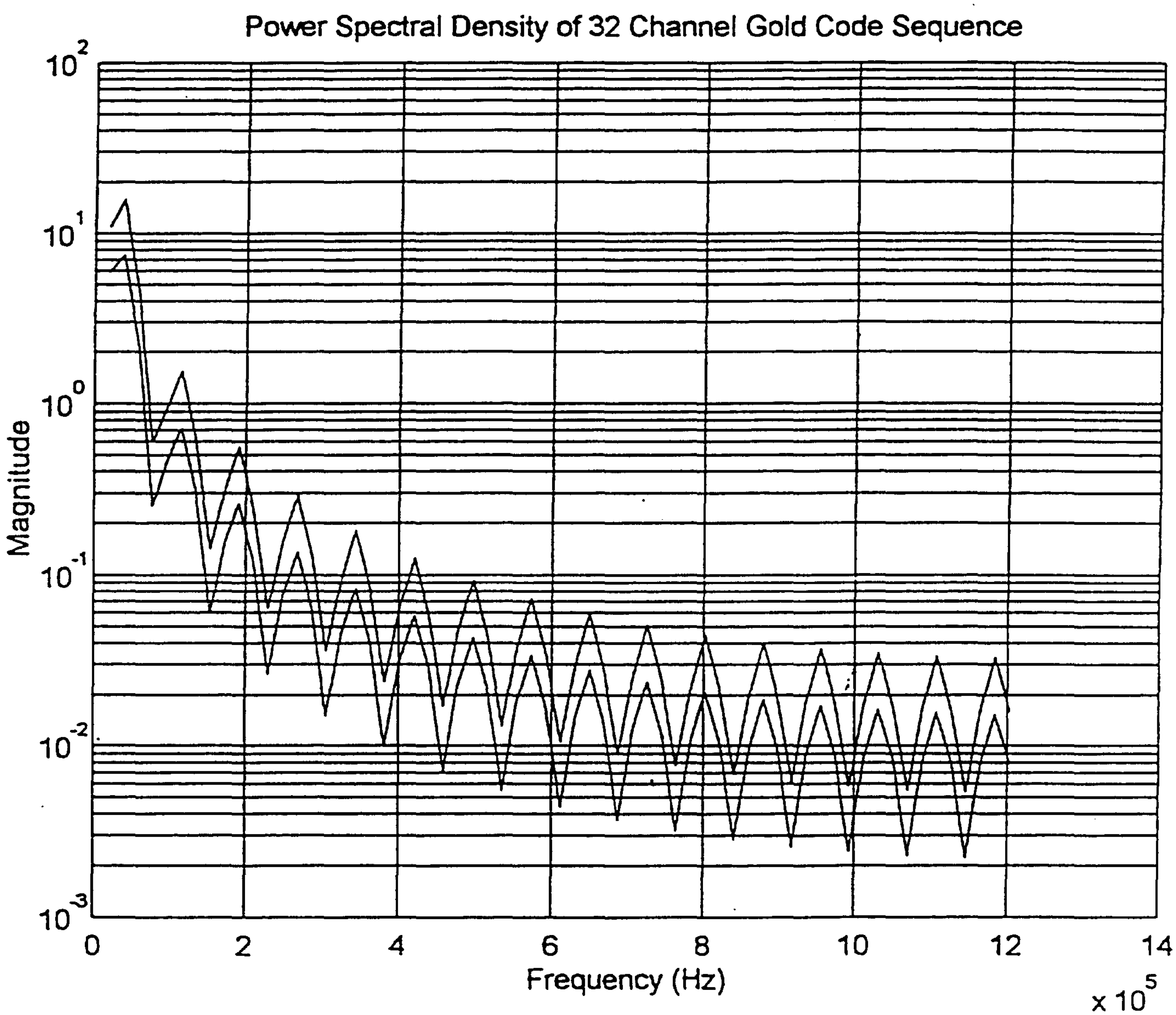

Figure 4. Baseband Spectrum Of NRZ Gold Code Sequence With Overlay Of 32 Signals, No Filtering.

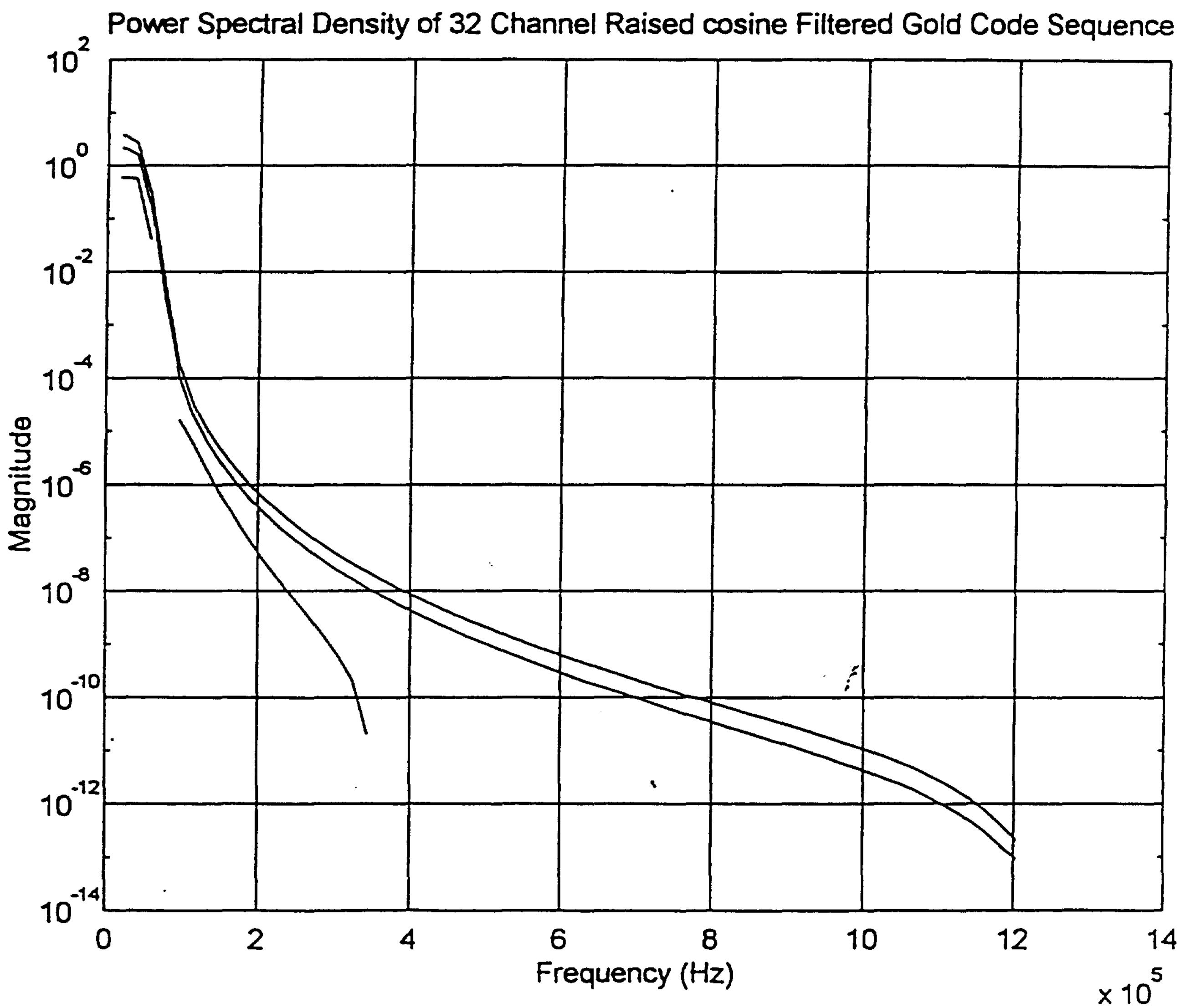

Figure 5. Baseband Spectrum Of NRZ Gold Code Sequence With 32 Signal Overlay, Beta = 1.0 Raised Cosine Filtering.